complexity. In addition, the use of a wideband DS signal offers the opportunity to use techniques not available with narrowband signals to mitigate the effects of multipath. The purpose of this subsection is to provide a brief summary of those techniques as they apply to the performance of wide-area locating/messaging systems.

1. The Basic Correlator Function

As noted above, the locating signal pulse for a wideband pulse-ranging system consists of a PN sequence with a specific "chip" pattern that is known to the receiver. At the heart of the receiver is a correlator that is matched to the PN sequence. As the received pulse propagates through the correlator, a correlation peak occurs when the PN sequence pattern matches the correlator pattern; i.e., when synchronization occurs. The TOA estimate is based on the time at which correlation peak occurs (or when the correlator output signal crosses some threshold). The rise time of the correlation peak is proportional to the chip rate and hence to the bandwidth of the transmitted pulse, and the uncertainty in the TOA estimate in the absence of multipath (for a given signal-to-noise ratio at the correlator output) is inversely proportional to the rise time. The uncertainty therefore varies inversely with the bandwidth of the ranging waveform, as indicated by the Cramer-Rao bound.

By definition, T_{RP} is the length of time over which the correlation occurs; that is, if T_C is the duration of a single chip and N_C is the length of the correlator in chips, then $E_{RP} = N_C T_C$. If n repetitions of the PN sequence are used and the individual TOA estimates are averaged, the resulting TOA estimation error is the same as if a single sequence of length $nN_C T_C = nT_{RP}$ were used *as long as* $E_{RP}/N_0 > \chi_{RP}$. If $E_{RP}/N_0 < \chi_{RP}$, even if nE_{RP}/N_0 (the total energy in n locating pulses) exceeds χ_{RP}, the receiver will be below threshold and will no longer perform according to (2). Thus, in evaluating system performance, it is important to distinguish between the duration of an individual locating pulse and the total duration of the burst *sequence* (n pulses). From an implementation perspective, there is a tradeoff here. While it is desirable to make the individual pulses as long as possible to maximize E_{RP}/N_0 and the "jamming margin" discussed earlier, there are practical limits on correlator length if SAW correlators are used (the required correlator length in millimeters is $T_{RP}(\mu\text{sec})/3$), or if information (data) signals are to be embedded in the ranging pulse (see below).

2. E_b/N_0 Threshold Reduction with Orthogonal Signaling

As observed above, pure DS spreading (simple modulation of the the data onto a PN sequence) for data communication does not reduce the E_b/N_0 threshold χ_b. Therefore, spreading alone does not affect performance against thermal noise, because the receiver front-end noise is increased by the same factor as the bandwidth, and this noise increase is exactly cancelled out by the post-correlation improvement in signal-to-noise ratio (the processing gain), resulting in no net change in the pre-detection signal-to-noise ratio. However, by spreading the signal in a more methodical manner, χ_b *can* be reduced. The most familiar example is the use of forward error correction (FEC), or "channel coding." With a rate 1/2 convolutional or block code, for example, the bandwidth is doubled (assuming the data rate and modulation format are fixed), and χ_b can be reduced several dB. The exact amount of the reduction depends on a number of factors, including the required BER and the type of code used, and the decoder implementation (e.g., "soft" vs. "hard" decision, etc.). However, FEC does not require a wideband signal and in fact is used in all of the new digital cellular radio formats, wideband and narrowband alike.

For wideband signals, "orthogonal signaling" can also be used. With this technique, each group of k information bits is mapped to one of $M = 2^k$ orthogonal waveforms. Orthogonality may be achieved in various ways, but for purposes of this discussion, the specific technique of interest is the use of M orthogonal PN sequences.

To understand the application of this technique to a locating/messaging system, suppose that k bits are to be embedded in each ranging pulse, which is assumed to be a single PN sequence of N_C chips. One possible implementation is to evenly divide the chips among the k bits so that each bit is modulated onto N_C/k chips (it is assumed here that $N_C >> k$). The problem with this is that there are now 2^k different possible incoming PN sequences, and the receiver can only correlate over N_C/k chips at a time. Thus, the jamming margins for both data communication and ranging are reduced by a factor of k, or $10\log k$ dB.

Another possible approach is to use 2^k different correlators, each matched to one of the k-bit patterns modulated onto the PN sequence. If the resulting 2^k sequences are orthogonal (which is possible if $2^k \leq N_C$), then this implementation is a form of "M-ary" orthogonal signaling. At the receiver, the correlator with the highest output would be assumed to be matched to the transmitted sequence, thereby recovering the transmitted data. That same correlator also would be used to estimate the pulse arrival time.

Compared to using an entire N_C-chip PN sequence to transmit each data bit using antipodal signaling (e.g., a bipolar baseband waveform and phase-shift keying with coherent detection), this approach reduces the processing gain by a factor of $10 \log k$ dB, but if 2^k orthogonal sequences are used, the E_b/N_0 threshold χ_b also is reduced. Fig. 3 shows the bit error probability vs. E_b/N_0 for orthogonal waveforms with several different values of k.

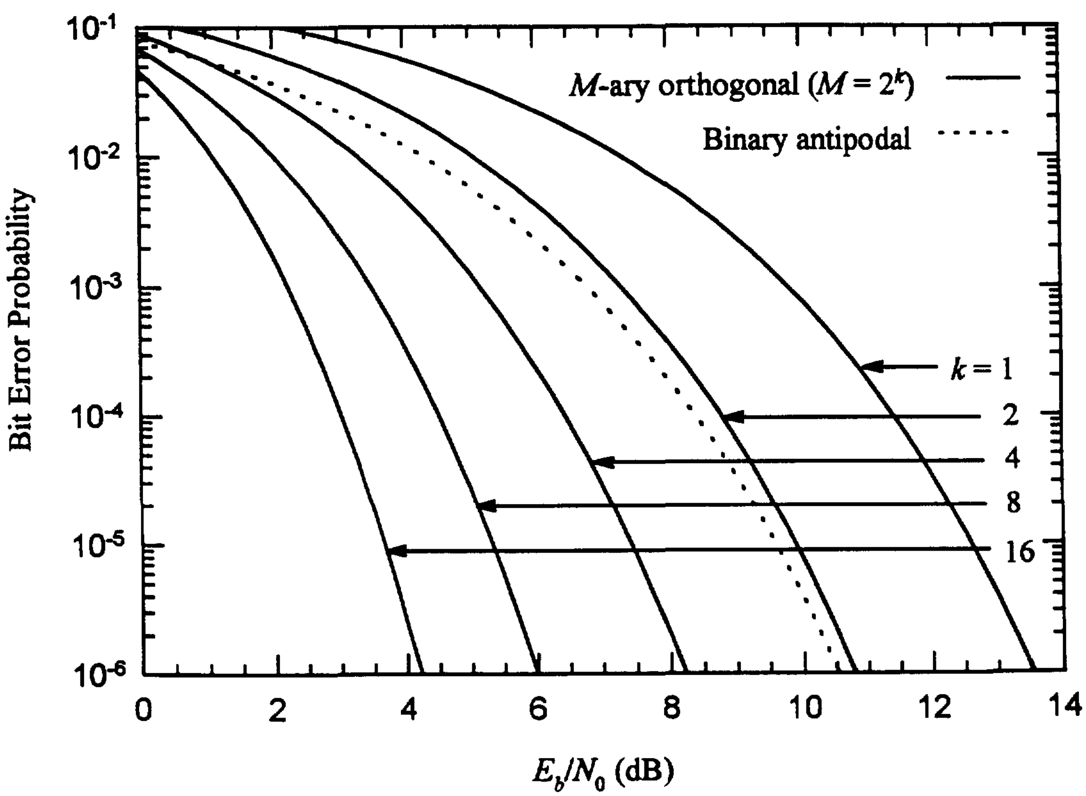

Figure 3. *Bit error probability vs. E_b/N_0 for orthogonal signals with coherent detection.*

These curves were computed using the method of Viterbi.* Also shown for

* A. J. Viterbi, *Principles of Coherent Communication*, New York: McGraw-Hill, 1966, section 8.3.

comparison is the "binary antipodal" case; i.e., a single bit per ranging pulse with binary PSK and coherent detection. As would be expected, the binary orthogonal case ($k = 1$) is 3 dB worse than the antipodal case.

For purposes of the particular problem being considered here, these curves are somewhat misleading, because if the ranging pulse duration is held constant, E_b/N_0 will decrease as k increases ($E_b = E_{RP}/k$). To illustrate, for $k = 4$, χ_b is reduced by about 5.3 dB compared to the $k = 1$ case (two orthogonal sequences), or about 2.3 dB less than if antipodal signaling is used. Overall, the jamming margin therefore is reduced by 6 - 2.3 = 3.7 dB for $k = 4$ (16-ary orthogonal signaling). The net results are (1) a four-fold increase in data rate; (2) a 6 dB reduction in the processing gain; (3) a 2.3 dB reduction (improvement) in χ_b; (4) a 3.7 dB reduction the jamming margin for data communication; and (5) an increase in the receiver complexity (effectively sixteen correlators instead of one). The net effect of increasing the number of bits per ranging pulse can be seen directly in Fig. 4, which shows the bit error probability vs. E_{RP}/N_0 rather than E_b/N_0. As can be seen, the E_{RP}/N_0 "threshold" required to achieve a given bit error rate increases with k. In fact, for $k = 2$, Fig. 4 suggests that two sequential antipodal signals (cutting E_{RP} by 3 dB) would provide better performance than quaternary orthogonal signaling. For further increases in k, orthogonal signaling is superior. For example, increasing k from 2 to 4 increases the E_{RP}/N_0 threshold by only about half a dB.

Finally, Fig. 5 shows the "symbol" error probability, denoted by P_S, vs. E_{RP}/N_0. Each block of k bits is mapped into one of M symbols, so detection actually occurs on symbols rather than bits. P_S is the probability that the decision circuitry selects a symbol other than that which was transmitted. As can be seen, P_S and the bit error probability differ very little for a given E_{RP}/N_0. What is noteworthy in the context of this discussion, however, is the fact that as k increases, the receiver requires an increasingly high E_{RP} to achieve a given P_S. It seems reasonable to assume that if the reception is sufficiently corrupted that a symbol error occurs, the TOA estimate also is corrupted. Thus, there appears to be a jamming margin penalty for the ranging application if multiple information bits are embedded in the ranging pulse. From Fig. 5, this penalty appears to be about 4 dB for $k = 4$ and 6 dB for $k = 8$ bits per pulse, compared to $k = 1$ bit per pulse using antipodal signaling.

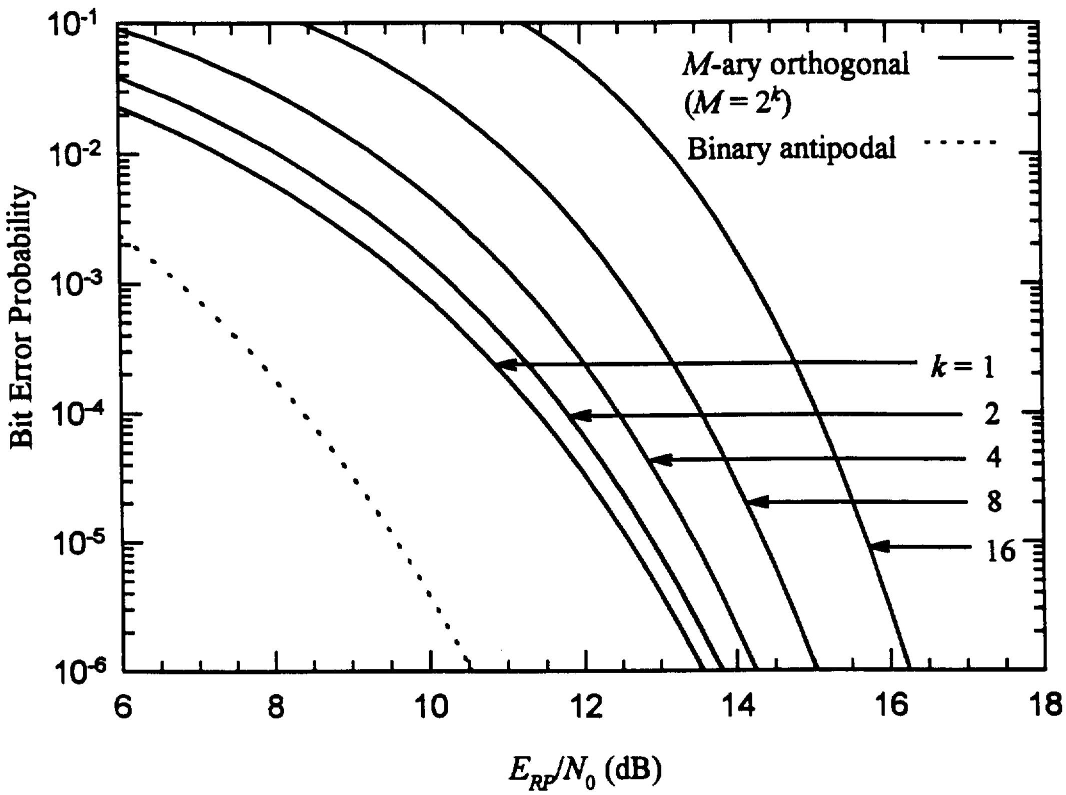

Figure 4. *Bit error probability vs.* E_{RP}/N_0 *for orthogonal signals with coherent detection.*

Finally, it should be noted that for a given value of k, biorthogonal signaling can be used to reduce the number of required correlators by a factor of 2 compared to orthogonal signaling. With biorthogonal signaling, 2^{k-1} orthogonal PN sequences are used, and each sequence is used to represent two k-bit sequences using antipodal signaling (conceptually, the demodulated output of each correlator is bipolar). On the basis of bit error probability, the performance of orthogonal and biorthogonal signaling is essentially the same; with respect to symbol error probability, biorthogonal signaling is superior for small k (3 dB better for k = 1).*

* See A. J. Viterbi, "On Coded Phase-Coherent Communication," *IRE Transactions on Space Electronics and Telemetry,* vol. 7, pp. 3-12, March 1961, Figs. 5,6,10, and 11.

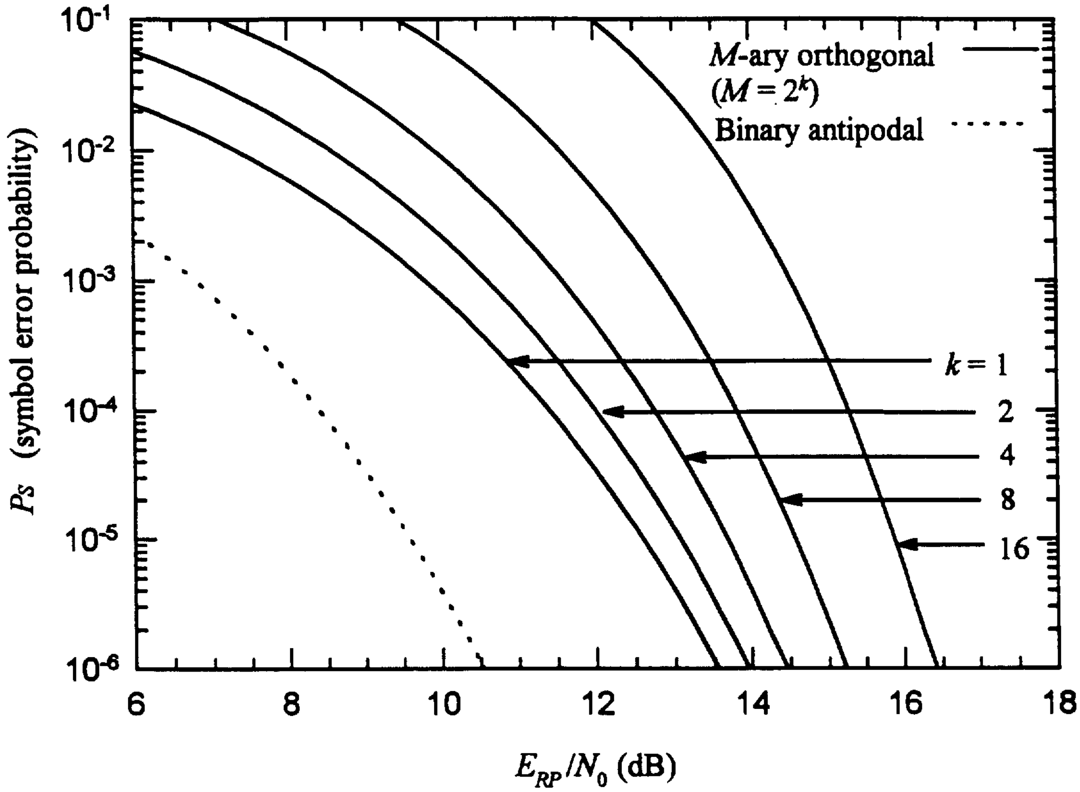

Figure 5. *Symbol error probability vs. E_{RP}/N_0 for orthogonal signals with coherent detection.*

3. Multipath Diversity with Multiple Correlators

Mobile communication in the multipath channel is a fairly complicated topic, and only a very brief high-level summary of pertinent aspects will be given here. There are a number of excellent books and articles that provide detailed treatments.*

The key parameters of a multipath channel are the delay spread and the Doppler. The delay spread is the time interval over which significant energy from multipath "echoes" arrive at the receiver, and it depends on the geometry of the environment. Environments with far-away "reflectors" of radio energy (e.g., a mountain range) will

* For a detailed discussion of the multipath channel for mobile radio, see W. C. Jakes, Jr. (ed.), *Microwave Mobile Communications,* New York: Wiley, 1974, chapter 1. For an extensive discussion of digital communication in the multipath channel, see Proakis, *op. cit.*, chapters 7 and 8.

tend to have large delay spreads. Environments such as the inside of an office building will tend to have small delay spreads because the major reflectors will be relatively near either the transmitter or the receiver. The "coherence bandwidth" varies inversely with the delay spread, and is a measure of the bandwidth over which the channel frequency response is likely to be relatively "flat." Doppler is a measure of how fast the channel impulse response (or equivalently, the transfer function in the frequency domain) is changing. For the mobile radio channel, the Doppler is due mainly to the motion of the vehicle. The "coherence time" is inversely related to the Doppler, and is a measure of the time interval over which the channel frequency response is relatively invariant.

The effect of the multipath channel on a digital communication system depends on the relationship between the system parameters and the delay/Doppler characteristics of the channel. For a signal that is narrowband relative to the coherence bandwidth, multipath causes quasi-periodic "flat fading" of the signal as the vehicle moves, meaning that the received signal power changes but the channel frequency response over the signal bandwidth is essentially flat. If the signal bandwidth exceeds the coherence bandwidth of the channel, the frequency response of the channel will exhibit noticeable variation across the signal bandwidth (i.e., the channel is frequency selective over the signal bandwidth). In the time domain, this means that the bit duration is not significantly greater than the delay spread. The result is intersymbol interference (ISI), and the usual remedy is an adaptive equalizer for a non-spread spectrum system.

The classical model for flat fading in the macrocell environment, in which a line-of-sight path often does not exist, is "Rayleigh fading," which means that the signal envelope (amplitude) is modeled as a Rayleigh-distributed random variable. Rayleigh fading causes large variations in the received signal power (typically on the order of 10 to 20 dB), which requires that extra "margin" be added to the link budget to assure a low probability that the signal fades below its threshold. One technique for combatting multipath is space diversity, which uses multiple antennas separated in space. The receiver would either select the antenna with the strongest signal ("selection diversity") or might combine the signals from the antennas in some way. Space diversity relies on the fact that in severe multipath (energy arriving from all 360° of azimuth), fades are quasi-periodic with minima an average of a half-wavelength apart. Thus, if two antennas are separated by a quarter-wavelength, it is statistically unlikely that both with fade simultaneously. The use of diversity reduces the amount of margin required to accommodate multipath fading.

Diversity also can be achieved by means other than multiple antennas. Time diversity can be used, providing the time span over which redundant information is transmitted exceeds the channel coherence time. Digital cellular radio air interfaces use a form of time diversity via forward error correction and interleaving of code symbols. Similarly, the use of frequency diversity requires that the redundant information be transmitted over a frequency span that exceeds the coherence bandwidth. Both frequency hopping and direct sequence modulation offer opportunities for forms of frequency diversity. With a direct sequence system that is wideband relative to the channel coherence bandwidth, a "rake" receiver architecture can be used, whereby multiple correlators are used to resolve the received signal into multiple path "clusters," which are essentially separate, independently-faded propagation paths. The correlator outputs then can be combined as diversity branches, reducing the required multipath margin. If the bandwidth of a direct sequence signal is sufficiently wide relative to the delay spread of the channel (i.e., a bandwidth on the order of 10 MHz for the outdoor mobile radio channel), even a single correlator can significantly reduce the received signal power variations due to multipath.*

4. Summary

The use of a wideband direct sequence spread spectrum signal provides the opportunity for correlator-based receiver techniques which can improve performance in two ways: (1) by reducing the E_b/N_0 threshold χ_b using orthogonal signaling and (2) by reducing the variations in received signal power due to multipath. Full use of these techniques can deliver a significant improvement in the link budget (which translates to a reduction in the necessary received signal power), although they result in a more complex and costly receiver. For purposes of comparing wideband and

* For details, see D. L. Noneaker and M. B. Pursley, "On the Chip Rate of CDMA Systems with Doubly Selective Fading and Rake Reception," *IEEE Journal on Selected Areas in Communications*, vol. 12, pp. 853-861, June, 1994.

narrowband forward links (in the next section), it will be assumed that the wideband link takes advantage of these techniques.

V. INTERFERENCE-LIMITED PERFORMANCE

It is assumed here that the effect of a cochannel interfering signal on the AVM/LMS receiver, with respect to both the locating and the messaging functions, will be the same as the effect of Gaussian noise of the same power. Due to the randomizing effect of the correlator on the interfering signal, this is a reasonable assumption for the present purpose. In terms of the parameters already introduced, this means that if I is the interference power, and it is substantially greater than the thermal noise power, then the effective noise spectral power density (from the perspective of the detector) is $N_0 = I/B$.

It also is assumed here that interfering transmitters are distributed over area in a random fashion, and that the average density (transmitters/km^2) active at a given time within a given bandwidth B is proportional to B (i.e., transmitters are also uniformly randomly-distributed in frequency).

A. Blocking Probability

Of interest is the "blocking probability," which is the probability that E_{RP}/N_0 falls below its threshold χ_{RP} (or for data communication, the probability that $E_b/N_0 < \chi_b$). As shown in the Appendix, the blocking probability can be expressed as a function of the average number of active interfering (in-band) transmitters within the service area of the AVM/LMS base station receiver (N) and the "near-far ratio" (*NFR*).* The *NFR* is defined as follows. If r_1 is the distance from the base station to the AVM/LMS mobile (transmitting the desired signal) and r_2 is the distance from the base station to the interfering transmitter, then blocking can occur if $r_2 \leq r_1/NFR$.

* The *NFR* definition was introduced in PR Docket 93-61 in "Interference Analysis of Part 15 Devices and LMS Wideband Systems - Initial Calculations," G. K. Smith, March 8, 1994, filed as Annex 2 with MobileVision's Comments on the *Ex Parte* presentations (March 15, 1994).

The blocking probability can be expressed in general as:

$$P_b = f\left(\frac{N}{(NFR)^2}\right), \tag{10}$$

where $f(\cdot)$ is a monotonically increasing function of its argument and has the properties that $\lim_{\alpha \to 0} f(\alpha) = 0$ and $\lim_{\alpha \to \infty} f(\alpha) = 1$. When only a single interference source is considered, and the AVM/LMS mobile randomly-positioned somewhere in its service area, the blocking probability is:

$$P_b = 1 - \frac{1}{\alpha}\left(1 - e^{-\alpha}\right), \tag{11}$$

where $\alpha = N/(NFR)^2$, which represents the average number of active transmitters near enough to the base station to cause interference when the mobile is at the coverage edge. The blocking probability for a mobile somewhere on the coverage edge is

$$P_b = 1 - e^{-\alpha}. \tag{12}$$

Eqs. (11) and (12) are somewhat optimistic because they consider only a single interference source. As shown in the Appendix, if the possibility of combined interference from multiple sources is considered, and path loss varies as r^{γ}, the blocking probability at the coverage edge is:

$$P_b(\alpha) = \frac{1}{\pi}\sum_{k=1}^{\infty} \frac{\Gamma(2k/\gamma)}{k!}\left[\alpha\Gamma(1-2/\gamma)\right]^k \sin k\pi(1-2/\gamma), \quad \alpha \geq 0, \tag{13a}$$

where $\Gamma(\cdot)$ is the Gamma function.* For the special case of $\gamma = 4$ (13a) reduces to:

$$P_b = \mathrm{erf}\left(\frac{\alpha\sqrt{\pi}}{2}\right), \tag{13b}$$

where erf$(\cdot)$ is the error function defined as $\mathrm{erf}(x) \equiv \frac{2}{\sqrt{\pi}} \int_0^x e^{-\xi^2}\, d\xi$.

The multiple-interferer blocking probability taken over the entire service area of the AVM/LMS base station is

$$P_b(\alpha) = \frac{1}{\alpha} \int_0^\alpha P_b(\xi) d\xi, \tag{14}$$

which can be evaluated numerically and is clearly a function of α. Fig. 6 shows P_b vs. $\alpha = N/(NFR)^2$ using (11) to (14). Note that for $P_b \leq 0.1$, the blocking probability over the service area can be closely approximated as $P_b \simeq N/2(NFR)^2$, and at coverage edge as $P_b \simeq N/(NFR)^2$. These approximations apply to both the single-interferer and multiple-interferer models.

B. Blocking Probability vs. Bandwidth

The *NFR* depends on the jamming margin of the AVM/LMS system as well as the transmitted power of the interference sources and the AVM/LMS mobile. The jamming margin depends on the processing gain and the E_{RP}/N_0 threshold, and the processing gain in turn depends on the system bandwidth. Assuming that path loss

* See chapter 6 of M. Abramowitz and I. E. Stegun, *Handbook of Mathematical Functions,* U. S. Dept. of Commerce, Nat. Bur. Stds., Dec., 1972.

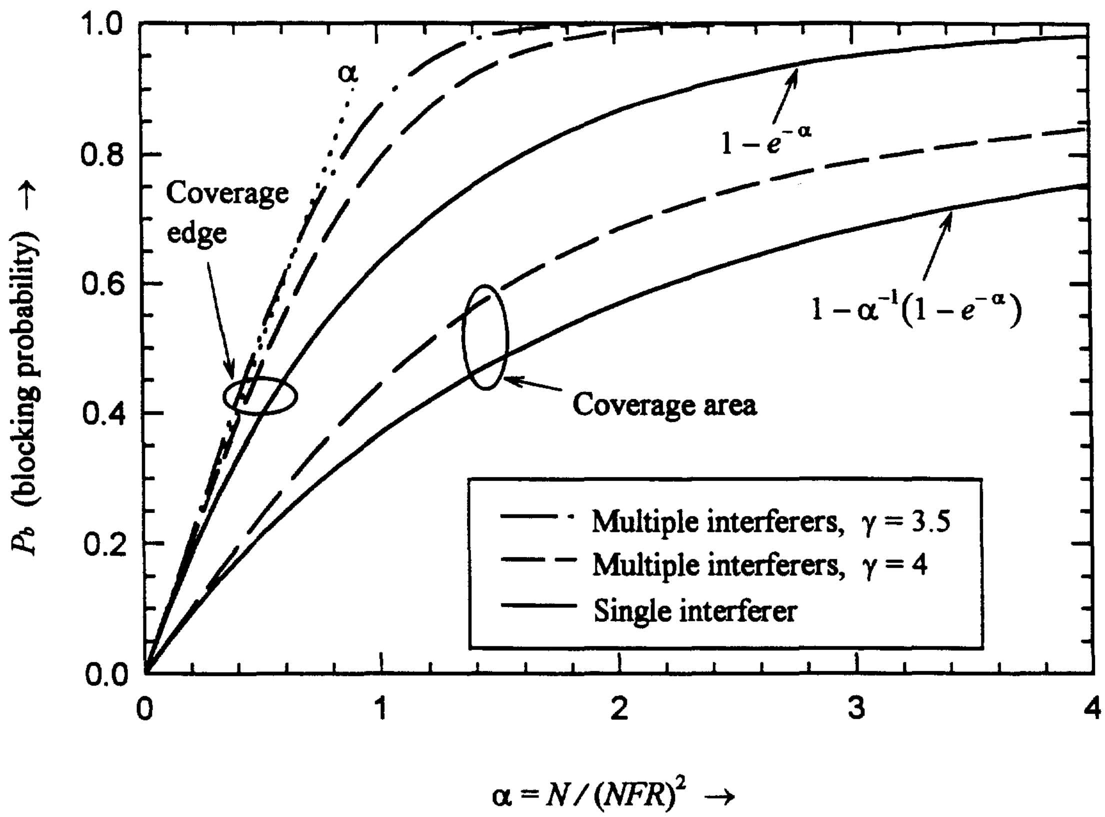

Figure 6. *Blocking probability for single-interferer and multiple-interferer models.*

over distance d varies as d^{γ}, the *NFR* can be expressed as:

$$NFR = (\tau B)^{1/\gamma}, \tag{15}$$

where the constant τ is:

$$\tau = \frac{P_D T_b}{P_I \chi_b} \quad \text{for data} \tag{16a}$$

$$\tau = \frac{P_D T_{RP}}{P_I \chi_{RP}} \quad \text{for ranging.} \tag{16b}$$

P_D and P_I represent the power transmitted by the desired and interfering transmitters. As before, T_b and T_{RP} are the durations of a data bit and of a ranging pulse, and χ_b and χ_{RP} are the thresholds for E_b/N_0 and E_{RP}/N_0, respectively.

Interfering transmitters are assumed to be randomly-distributed in position and frequency, so the average number of active transmitters within the service area and the bandwidth of the AVM/LMS system will be proportional to the bandwidth and the service area. Thus,

$$N = \kappa AB \,, \tag{17}$$

where κ is the average number of interfering transmitters per km^2 per MHz and A is the size of the AVM/LMS base station service area in km^2.

Combining (10), (15), and (17) gives:

$$P_b = f\left(\frac{\kappa AB}{r^{2/\gamma} B^{2/\gamma}}\right) = f\left(\frac{\kappa A}{r^{2/\gamma}} \cdot B^{1-2/\gamma}\right) . \tag{18}$$

It is clear from (18) that for $\gamma > 2$, the argument of $f(\cdot)$ increases with B, and hence the blocking probability increases as the bandwidth of the AVM/LMS system increases.* It is therefore a losing proposition to expand bandwidth to achieve additional processing gain with which to overcome the effects of interference sources that are randomly-distributed in position and frequency.†

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* It should be noted that $\gamma > 2$ is a condition for systematic frequency reuse, and γ is often taken as 4 for modeling the mobile radio environment.

† In a totally "homogeneous" environment such as a cellular system the conclusion is different, because the transmitted power levels of all the mobiles are subject to the control of a coordinated system. Received power levels then can be managed to eliminate the "near-far" problem.

For the case of $\gamma = 4$, (18) becomes:

$$P_b = f\left[\kappa A \sqrt{B/\tau}\right] . \tag{19}$$

Maintaining P_b constant as B is increased therefore requires some combination of (1) reduction in service area; (2) increasing the transmit power; (3) increasing the bit or ranging pulse duration (which will decrease "capacity"); and (4) reducing χ_{RP} or χ_b.

This multi-parameter tradeoff is easily quantified by considering two systems with bandwidths B_1 and B_2, and plugging (16) into (18). For example, assuming that $\gamma = 4$ and that the two systems have the same blocking probability, this yields:

$$\frac{A_1^2}{A_2^2} \cdot \frac{P_2 T_2/\chi_2}{P_1 T_1/\chi_1} = \frac{B_2}{B_1} , \tag{20}$$

where, for the system indicated by the subscript:

A	is the base site service area.
P	is the effective radiated power.
T	is the duration of either an information bit or a ranging pulse, depending on which case is being considered.
χ	is the threshold for either E_b/N_0 or E_{RP}/N_0, depending on which case is being considered.
B	is the signal bandwidth.

C. Implications for Wide-Area AVM/LMS Systems

It is worthwhile to consider in some detail the implications of (20) on wide-area locating systems operating in the 902-928 MHz band. For the reverse (mobile-to-base) link, which is necessarily fairly wide-band because of its pulse-ranging function, the significance of (20) is relatively mild, because the practical range in the ratio B_2/B_1 is fairly limited. For example, when comparing a reverse link with $B = 16\,\text{MHz}$ to one with $B = 4\,\text{MHz}$, $B_2/B_1 = 4$, which from (20) can be accommodated by a halving of the service area, or a 6 dB increase in transmit power, or a fourfold increase in T_{RP} (which translates to a quartering of capacity). It is interesting to note that assuming A, P, and χ are fixed, (20) shows that for $\gamma = 4$, capacity actually varies *inversely* with bandwidth, if the dominant impairment is interference from randomly-positioned transmitters. This is in direct contradiction to Pinpoint's claim that locating capacity increases as the "bandwidth cubed" when performance is limited by narrowband cochannel interference.* Pinpoint based this conclusion on the Cramer-Rao bound argument already discussed, plus the fact that the effective post-correlation spectral power density of the noise (N_0) varies inversely with the bandwidth of the desired signal (which is true). Pinpoint's analysis was incomplete, however, due to (1) a failure to account for the receiver threshold, as already discussed, and (2) a failure to account for the fact that as bandwidth increases, so does the probability that there will be an in-band transmitter near enough to the receiver to cause interference. Pinpoint hypothesized an in-band interference source of fixed power, and hence ignored the inherently statistical nature of the interference problem. In a more recent filing, Pinpoint acknowledges the statistical nature of the interference problem, stating as one of its headings that "Interference is a Matter of Spatial and Temporal Probability,"† but has failed to correct its previous claims of the bandwidth advantage in the presence of cochannel interference.

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* Louis Jandrell (June 29, 1993), *op. cit.*, p. 9.

† David E. Hilliard, Attorney for Pinpoint Communications, Inc., in a letter to Mr. Ralph Haller in association with PR Docket 93-61, June 27, 1994, p. 3.

Notwithstanding the fact that a widening of the reverse link beyond 4 MHz moves the "power vs. capacity vs. service area" tradeoff to a less favorable point in the design space, it may be justifiable on the basis of improved accuracy, and is a design decision that perhaps is best left to the service provider. For the forward link, however, the conclusion is quite different, because the forward link is essentially a wide-area paging channel and does not need to perform a ranging function. Thus, there is no inherent reason to use a wideband forward link.

As an example of the impact of using a wideband forward link, consider Pinpoint's proposed ARRAY™ system. The bit rate is on the order of 300 kb/s* and the average forward link duty cycle is on the order of 3 percent,† so the average net throughput is about 9 kb/s. Even allowing for overhead associated with symbol and frame synchronization, error control, and guard bands, a 9 kb/s data throughput should be easily accommodated in a channel 25 kHz wide. Given that the blocking probability and service areas are fixed, (20) indicates that PT/χ can be reduced by a factor of 640, or 28 dB by using a 25 kHz channel compared to a 16 MHz channel. Further, since the bit duration T will increase by a factor of $1/0.03 = 33$, the total difference in P/χ is $28 + 10 \log 33 \simeq 43\,\mathrm{dB}$. If the wideband forward link takes full advantage of the wideband receiver techniques discussed in Section IV, the narrowband (25 kHz) channel will require substantially more margin for multipath than the wideband channel (assuming no diversity is used in the mobile unit's receiver), and will have a higher value of χ_b in the absence of multipath. Assuming that for the 16 MHz channel, the use of orthogonal signaling and multiple correlator diversity reduces χ_b and the multipath margin by 8 dB and 15 dB, respectively, the net reduction is 23 dB and the wideband link requires 20 dB, rather than 43 dB, more transmit power. Thus a forward link that would require 100 watts ERP using a 25 kHz channel would need to transmit 10 kW ERP using a 16 MHz channel. This calculation is remarkably consistent with Pinpoint's contention that the allowed forward link power should be 5 kW.

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* Pinpoint Communications Inc., *Ex Parte* Communication in PR Docket 93-61, May 11, 1994, p. 3.

† David E. Hilliard, *op. cit.*, p. 5.

Pinpoint's reason for this is explained in its June 29, 1993 Comments:*

> However, in order to maintain reliable operation of the wide-area systems over the communications ranges needed for economic infrastructure deployment, typically greater than 5 miles average spacing between base stations, the base station power levels will need to be able to operate up to 5 kilowatt ERP in order to be able to ensure that the mobile's [*sic*] will be able to receive the base signals while near to local-area system noise/jamming sources.

In addition to the disadvantage in efficiency associated with the wideband forward link, the receiver (which is in the mobile unit) has been assumed to use the wideband receiver techniques discussed in Section IV to minimize multipath effects and χ_b. This will result in a receiver architecture that is more complex and costly than a receiver for a narrowband forward link. In this latter case (a 25 kHz channel), a simple frequency-shift keying (FSK) architecture with discriminator detection could be used, which is one of the simplest and least costly digital receiver architectures available.

Although the use of narrowband rather than wideband forward links is more efficient for the service provider and will likely simplify the receiver in the mobile unit somewhat, the larger benefits of using narrowband forward links become evident when the "total picture" is considered. A wideband forward link must be "time-shared" with the reverse links of the AVM/LMS service providers, so the use of the wideband forward link reduces reverse link capacity. In addition, the narrowband forward links are much less likely to cause irresolvable interference problems for other users of the band than are wideband forward links. Although the forward link is necessarily high-power, narrowband forward links located near the band edges would be relatively easy for frequency-agile systems to avoid. A wideband forward link, despite its intermittent nature and lower spectral power density, probably would be impossible for some users to totally avoid.

* Comments of Pinpoint Communications, Inc., on the NPRM in PR Docket 93-61, June 29, 1993, p. 29. See also p. 32 of Pinpoint's March 15, 1994 Comments on the *Ex Parte* presentations.

It is interesting to consider the effect of a 5 kW, 16 MHz forward link on other users in the band (e.g., Amateur radio operators and Part 15 devices). Assuming that the "victim" receiver has an effective IF bandwidth of 100 kHz* and an 8 dB overall noise figure, its thermal noise floor is -116 dBm. The effective radiated power from the 5 kW wideband forward link into the 100 kHz receiver is 45 dBm (obtained by multiplying the ratio 100 kHz/16 MHz by the 5 kW ERP). Therefore, the signal received by victim receiver from the AVM/LMS forward link is 20 dB above the noise floor if the path loss is 45+116-20 dB = 141 dB. Fig. 7 shows the median path loss based on the "Hata" model,† assuming that the AVM/LMS forward link transmit antenna is elevated 300 feet and the victim receive antenna is elevated 30 feet (i.e., an outdoor pole-mounted Part 15 device). As can be seen, the separation required for a median path loss of 141 dB depends on the environment, but for a suburban area, more than 20 miles of separation is required. For the "large city" and "medium/small city" environments, roughly 5 miles and 11 miles of separation would be needed, respectively.

For a receiver nearer the ground, the interference prospects are somewhat less severe, but still serious. Fig. 8 shows the median path loss if the receiver is 5 feet above the ground‡ (e.g., a wireless data terminal or a cordless telephone). A path loss of 141 dB requires separations of roughly 2.8 miles for the urban environment, about 5.7 miles for the suburban environment, and roughly 20 miles for the "open area" environment.

Even from these simple calculations it is clear that wideband high-power forward links are by no means benign with respect to their potential for interfering with other users of the band. Although the duty cycle of a given forward link transmitter may be relatively low, the interference radii are sufficiently large that a given victim receiver will likely be within range of multiple forward link transmitters, and if AVM/LMS

* A direct sequence spread spectrum system with a 1 MHz bandwidth and 10 dB of processing gain is equivalent to a non-spread system with a 100 kHz bandwidth for purposes of this calculation.

† M. Hata, "Empirical Formula for Propagation Loss in the Land Mobile Radio Services," *IEEE Transactions on Vehicular Technology*, vol. VT-29, no. 3, August, 1980, pp. 317-325.

‡ For a receiver 5 feet above the ground, the path losses given by the Hata model for the "Large city" and "Medium/small city" environments are the same, and are simply labeled "Urban" in Fig. 8.

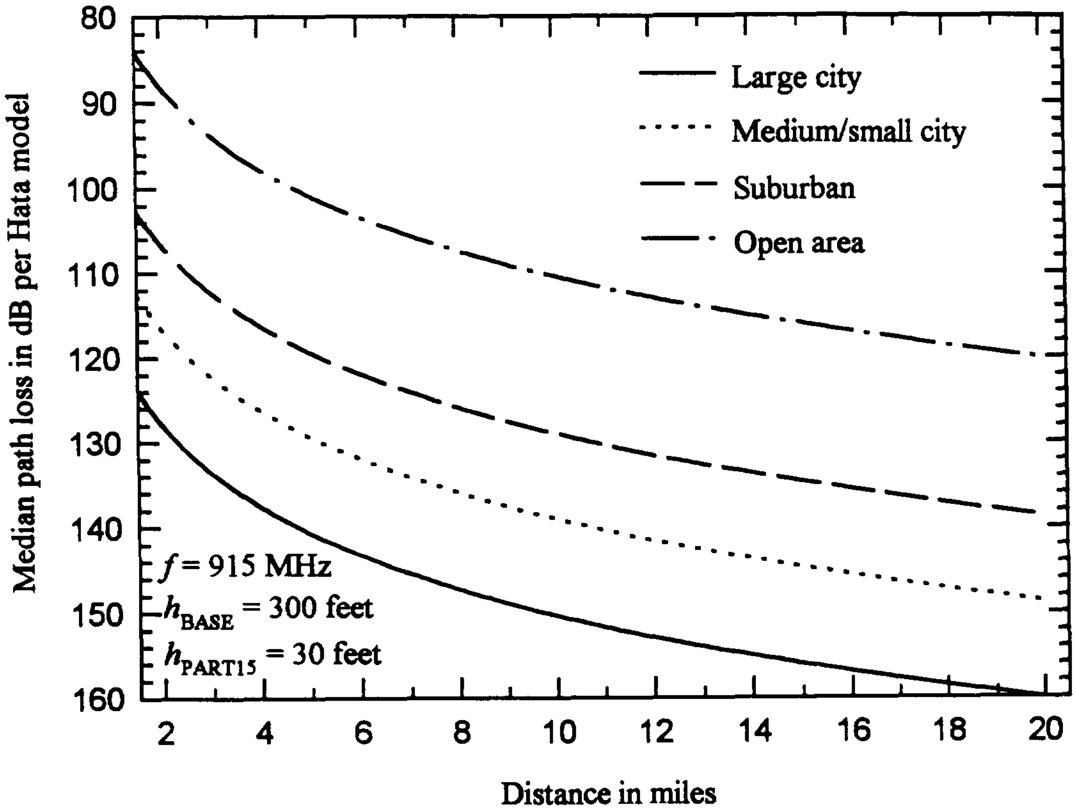

Figure 7. *Median path loss vs. distance for 915 MHz with 300-foot base and 30-foot receive antenna (per Hata).*

systems using wideband forward links are widely deployed, interference to other band users will become a chronic problem. Therefore, the use of narrowband forward links is much better suited to a band such as 902-928 MHz, which may be shared among other systems (including other AVM/LMS networks) providing a wide variety of services.

VI. CONCLUSIONS AND RECOMMENDATIONS

The main result of the analyses provided here is to dispel two major misconceptions about the effect of bandwidth on the performance of wide-area AVM/LMS systems. The two misconceptions are (1) that in the absence of cochannel interference, maximum locating capacity increases as the square of the bandwidth, and (2) that in the interference-limited case, locating capacity increases as the cube of the bandwidth. Related to this second idea is the misconception that in an environment with interfering transmitters that are randomly-distributed over area and in frequency, the

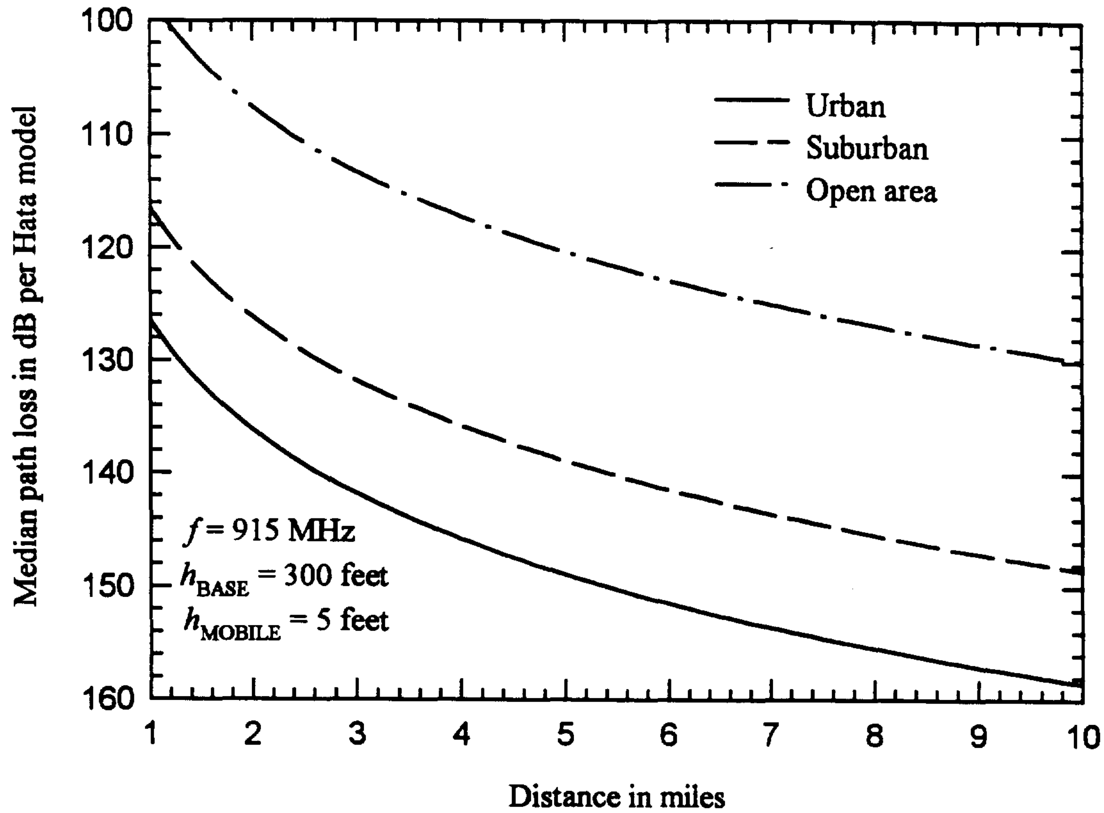

Figure 8. *Median path loss vs. distance for 915 MHz with 300-foot base and 5-foot receive antenna (per Hata).*

use of a wideband signal is somehow more efficient in terms of sharing spectrum with the other transmitters. This erroneous notion is summed up by a statement in a recent *Ex Parte* presentation by Pinpoint, which claims: "Much like the 'Steinbrecher Box,' Pinpoint's system uses wider bandwidth to increase the spectrum efficiency over that which can be achieved through simple, narrower bandwidth frequency division."*

It has been shown that in the absence of cochannel interference, multipath is the dominant limitation on locating accuracy, and increasing the reverse-link bandwidth beyond the 4 MHz minimum discussed in PR Docket 93-61 apparently improves the locating accuracy due to better multipath resolution but bears no relationship to

* Pinpoint Communications Inc., *Ex Parte* Communication in PR Docket 93-61, May 11, 1994, p. 11.

capacity (fixes per second). Rather, locating capacity is determined by a combination of the effective radiated power (ERP) from the mobile unit and the spatial density of the base stations. The large capacity (1000 or more fixes/second) claimed by Pinpoint for its proposed ARRAY™ system is due to the 40 watt mobile transmitters and the relatively close spacing between base stations, compared to Teletrac's 70 fix/second capacity using mobile units with roughly 1 watt ERP and a less dense network of base stations. Thus, in the interest of locating accuracy it may make sense to use a relatively wide-band reverse (mobile-to-base) link, but capacity will be independent of the reverse-link bandwidth, and will be determined by the reverse-link transmit power and the density of the base station deployment.

It also has been shown that the use of a wideband channel for communication in the presence of interference sources that are randomly-distributed in space and frequency is actually *less* efficient than the use of a narrowband channel. Simply stated, this is due to the fact that as the bandwidth increases, so does the probability that there will be an active transmitter near enough to the receiver to cause interference, even if the increase in "processing gain" with bandwidth is taken into account. For a median path loss that varies as distance to the fourth power (a common model for approximating the mobile radio environment), the data capacity actually varies *inversely* with the bandwidth, all other system parameters, including blocking probability, being fixed. If capacity is to be held constant, the transmitted power must be increased in proportion to bandwidth. While the use of wideband receiver techniques (rake-like multiple correlator receivers, orthogonal signaling) can reduce this inverse-bandwidth variation somewhat, it will not totally eliminate it, and increasing the bandwidth will still reduce capacity unless the transmitted power is increased.

This effect does not seem particularly significant for the reverse link, which requires a relatively large bandwidth because of its pulse-ranging function; as a practical matter, the benefits of increased accuracy from widening the reverse link beyond 4 MHz may outweigh the disadvantages of the modest power increase required to maintain the capacity and "blocking probability" constant. However, the implications on the forward link *are* significant, because the forward link is essentially a paging channel, and there is no inherent need for it to be wideband. Based on the required data rate and duty cycle cited by Pinpoint, the average throughput required for the forward link could be accommodated in a channel 25 kHz wide. Even considering the forward link in isolation, this would be a more spectrum-efficient solution for the AVM/LMS service provider, and would reduce the complexity of the mobile receiver. Perhaps a more important benefit, however, of using narrowband rather than wideband forward links, is that the availability of the spectrum for *reverse-link* operations of multiple AVM/LMS service providers would be improved, and the potential for interference with other users of the band would be greatly reduced.

These conclusions suggest the basis for a set of FCC rules which would promote coexistence among multiple AVM/LMS systems, and between AVM/LMS systems and other users, in the 902-928 MHz band:

- Wideband, high-power forward links should not be allowed. Multiple narrowband (e.g., 25 kHz) channels near the band edges should be designated for forward links. Each wide-area AVM/LMS service provider could be granted exclusive use of one or several forward channels in a given territory.
- The availability of a relatively large block of spectrum (perhaps 10 to 16 MHz) may be justified for the reverse link on the basis of locating accuracy in the presence of multipath, with limits on the mobile transmit power and duty cycle. If necessary, a protocol for "time-sharing" of the reverse link spectrum among multiple service providers coexisting in a given area could be developed cooperatively by the wide-area AVM/LMS interests, possibly under the auspices of a standards forum such as the TIA (Telecommunications Industry Association).

This framework should minimize the potential for interference between AVM/LMS systems and other users of the band, while allowing AVM/LMS system designers a relatively high degree of design freedom.

APPENDIX

DERIVATION OF THE BLOCKING PROBABILITY WITH COMBINED INTERFERENCE FROM MULTIPLE TRANSMITTERS

The blocking probability is easily derived when the interference from only a single source is considered. In reality, however, the total interference power at any given instant of time will be the sum of the interference from multiple co-channel "undesired" transmitters. The analysis presented here accounts for the multiple-source aggregate interference.

A. Blocking Probability - General Expression

Assume a normalized distance scale such that the average density of interference sources transmitting within the band of interest at a given time is $1/\pi$ (interferers per unit normalized area). If the "victim" receiver (i.e., an AVM/LMS system base station) is at the center of a circle of normalized radius $\sqrt{K}$, the expected (average) number of interference sources within the circle is K. Assuming that interfering transmitters are randomly distributed over area in a uniform fashion, the *actual* number of active interfering transmitters within the circle at a given time can be modeled as a Poisson-distributed random variable **J** with discrete probability density function (pdf):

$$P_{\mathbf{J}}(j) = P\{\mathbf{J} = j\} = \frac{e^{-K} K^{j}}{j!}\,, \tag{A-1}$$

where the notation $P\{\cdot\}$ represents the probability of the indicated event. The power received at the base station from an interfering transmitter a normalized distance s away from it is proportional to $s^{-\gamma}$. Since the aim here is to study the statistics of the carrier-to-interference ratio, the proportionality constant can be set to 1, and the ratio of the effective radiated power levels for the desired and interfering signals can be subsumed into the normalized carrier (desired signal) power. If the interference power from the jth interfering transmitter is $\mathbf{z}_j = \mathbf{s}_j^{-\gamma}$, then the total interference from interfering transmitters within the circle of normalized radius $\sqrt{K}$ is:

$$Z_K = \sum_{j=1}^{J} \mathbf{z}_j . \tag{A-2}$$

With interferers that are randomly distributed over area, the pdf of $\mathbf{s}_j$ is:

$$f_{\mathbf{s}_j}(s) = \frac{2s}{K}, \quad 0 \leq s \leq \sqrt{K} . \tag{A-3}$$

Hence, the pdf of $\mathbf{z}_j$ is

$$f_{\mathbf{z}_j}(z) = \frac{2}{\gamma K} z^{-(\gamma+2)/\gamma}, \quad K^{-\gamma/2} \leq z \leq \infty . \tag{A-4}$$

If the desired transmitter is a normalized distance D from the base station, the normalized desired signal power is $aD^{-\gamma}$, where the parameter a accounts for the ratio of the effective radiated power (ERP) of the desired transmitter to that of an interfering transmitter. "Blocking" is assumed to occur if the carrier-to-interference ratio C/I falls below some threshold X. The blocking probability therefore is:

$$\begin{aligned} P_b &= P\left\{\frac{C}{I} < X\right\} = P\left\{\frac{aD^{-\gamma}}{\mathbf{Z}} < X\right\} \\ &= P\left\{\mathbf{Z} > \frac{aD^{-\gamma}}{X}\right\} = 1 - P\left\{\mathbf{Z} < \frac{aD^{-\gamma}}{X}\right\} \\ &= 1 - F_{\mathbf{Z}}\left(\frac{aD^{-\gamma}}{X}\right), \end{aligned} \tag{A-5}$$

where $\mathbf{Z} = \lim_{K \to \infty} \mathbf{Z}_K$ and $F_{\mathbf{Z}}(z) = P\{\mathbf{Z} < z\}$ is the cumulative distribution function (CDF) of the random variable **Z**.

Note that the expected number of interfering transmitters (N) within a circle of radius D is D^2, and that the "near-far ratio" (NFR) is:

$$NFR = (a/X)^{1/\gamma}. \tag{A-6}$$

Therefore:

$$\frac{aD^{-\gamma}}{X} = \left(\frac{N}{(NFR)^2}\right)^{-\gamma/2}. \tag{A-7}$$

Letting $\alpha = N/(NFR)^2$, (A-5) gives the blocking probability at the edge of the coverage area as:

$$P_b = 1 - F_{\mathbf{Z}}(\alpha^{-\gamma/2}). \tag{A-8}$$

Determining the blocking probability for the multiple-interferer case therefore requires finding the CDF of the aggregate interference power **Z**, which can be accomplished by determining the pdf of **Z** from its characteristic function (or moment-generating function) and integrating.

B. The Characteristic Function of the Aggregate Interference

The characteristic function of **Z** is:

$$\Phi_{\mathbf{Z}}(\omega) = E\left[e^{j\omega\mathbf{Z}}\right] = \int_0^{\infty} f_{\mathbf{Z}}(z) e^{j\omega z} dz, \tag{A-9}$$

which is the Fourier transform of $f_{\mathbf{Z}}(z)$, with $E[\cdot]$ denoting the expectation operator